ABSTRACT

This application report covers the design of a Sallen-Key highpass biquad. This design gives low component and op amp sensitivities. It also shows how to compensate for the op amp's bandwidth (pre-distortion) and parasitic capacitances. A design example illustrates this method. These biquads are also called KRC or VCVS [voltage-controlled, voltage-source].

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Introduction

Changes in component values over process, environment and time affect the performance of a filter. To achieve a greater production yield, the filter needs to be insensitive to these changes. This application report presents a design algorithm that results in low sensitivity to component variation. See reference [6] for information on evaluating the sensitivity performance of your filter.

To achieve the best production yield, the nominal filter design must also compensate for component and board parasitics. The components are pre-distorted (reference [5]) to compensate for the op amp bandwidth. This application report expands the pre-distortion method in reference [5] to include compensation for parasitic capacitances. This method is valid for either voltage-feedback or current-feedback op amps.

Parasitic Compensation

To pre-distort your filter components and compensate for parasitic capacitances:

1. Use the method in reference [5] to include the op amp’s effect on the filter response. The result is a transfer function of the same order whose coefficients include the op amp group delay ($\tau_{oa}$) evaluated at the passband edge frequency ($f_c$).

2. For all parasitic capacitances in parallel with capacitors:
   • Add the capacitors together
   • Simplify the resulting coefficients
   • Use the sum of time constants form for the coefficients when possible

3. For all parasitic capacitances in parallel with resistors:
   • Replace the resistor $R_x$ in the filter transfer function with the parallel equivalent of $R_x$ and $C_p$.
     $$R_x \leftarrow \frac{R_x}{1 + R_xC_p s} , \ s \rightarrow j \omega$$
   • Alter this impedance to a convenient form and simplify:
     • Do not create new terms (a coefficient times a new power of $s$) in the transfer function after simplifying
     • The most useful approximations are:
       $$\frac{R_x}{1 + R_xC_p s} \approx R_x \left( 1 - R_xC_p s \right)$$
       $$\approx R_x e^{-R_xC_p s}$$
   These approximations are valid when:
   $$\omega << \frac{1}{R_xC_p}$$
   • Convert $(1 + R_xC_p s)$ to the exponential form (a pure time delay) when it multiplies, or divides, the entire transfer function
   • Do not change the gain at $\omega \approx \omega_p$ in allpass sections
   • When simplifying, discard any terms that are products of the error terms ($k\tau_{oa}$ and $R_xC_p$); they are negligible
   • Use the sum of time constants form for the coefficients when possible

Use an op amp with adequate bandwidth ($f_{3dB}$) and slew rate (SR):

$$f_{3dB} \geq 10f_i$$  \hspace{1cm} (1)
$$SR > 5f_iV_{\text{peak}}$$ \hspace{1cm} (2)

where $f_i$ is the highest frequency in the passband of the filter, and $V_{\text{peak}}$ is the largest peak voltage. This increases the accuracy of the pre-distortion algorithm. It also reduces the filter’s sensitivity to op amp performance changes over temperature and process. Make sure the op amp is stable at a gain of $A_V = K$. 
3 KRC Highpass Biquad Design

The biquad shown in Figure 1 is a Sallen-Key highpass biquad. \( V_{\text{in}} \) needs to be a voltage source with low output impedance.

The transfer function is:

\[
\frac{V_{o}}{V_{\text{in}}} = \frac{H_{\infty} \left( \frac{1}{\omega_p^2} \right) s^2}{1 + \left( \frac{1}{\omega_p Q_p} \right) s + \left( \frac{1}{\omega_p^2} \right) s^2}
\]

\[
K = 1 + \frac{R_f}{R_g}
\]

\[
H_{\infty} = K \left( \frac{1}{\omega_p Q_p} \right) = R_2 C_1 + R_3 C_3 - R_4 C_3 (K - 1)
\]

\[
\frac{1}{\omega_p^2} = R_2 R_3 C_1 C_3
\]

![Figure 1. Highpass Biquad](image)

To achieve low sensitivities, use this design algorithm:

1. Partition the gain for good \( Q_p \) sensitivity and dynamic range performance:
   - Use a low noise amplifier before this biquad if you need a large gain
   - Select \( K \) with this empirical formula:
     \[
     K = \begin{cases} 
     1, & 0.1 \leq Q_p \leq 1.1 \\
     2.2 Q_p - 0.9, & 1.1 < Q_p < 5
     \end{cases}
     \]

     These values also reduce the op amp bandwidth’s impact on the filter response. This biquad’s sensitivities are too high when \( Q_p \geq 5 \)

2. Select an op amp with adequate bandwidth \( (f_{3dB}) \) and slew rate (SR):
   - \( f_{3dB} \geq f_c \cdot S_{3dB} \geq 10f_c \cdot S_R \geq 5f_c \cdot V_{\text{peak}} \)
   - where \( f_c \) is the highest signal frequency, \( f_c \) is the corner frequency of the filter, and \( V_{\text{peak}} \) is the largest peak voltage. Make sure the op amp is stable at a gain of \( A_v = K \).

3. For current-feedback op amps, use the recommended value of \( R_f \) for a gain of \( A_v = K \). For voltage-feedback op amps, select \( R_f \) for noise and distortion performance. Then set \( R_g \) for the correct gain:
   \[
   R_g = \frac{R_f}{(K - 1)}
   \]
4. Initialize the resistance level \( R = \sqrt{R_4R_5} \). Increasing \( R \) will:
   - Increase the output noise
   - Reduce the distortion
   - Improve the isolation between the op amp outputs and \( C_1 \) and \( C_3 \)
   - Make the parasitic capacitances a larger fraction of \( C_1 \) and \( C_3 \)

5. Initialize the capacitance level \( S^{H_p} \alpha_i \), and the component ratios
   \[
   \begin{align*}
   c^2 &= \frac{C_3}{C_1} \quad \text{and} \quad r^2 = \frac{R_5}{R_4} \\
   c^2 &= \frac{1}{(\alpha_0R)} \\
   c^2 &= 0.10 \\
   r^2 &= \max\left\{0.10, \frac{1 + \sqrt{1 + 4Q_p^2(1 + c^2)(K - 1)}}{2 \cdot Q_p \cdot (1 + c^2) / c}\right\}
   \end{align*}
   \]

6. Recalculate \( C^2 \) and initialize the capacitors:
   \[
   c^2 = \left(\frac{2 \cdot r \cdot Q_p}{1 + \sqrt{1 + 4Q_p^2(K - 1 - r^2)}}\right)^2
   \]
   \[
   C_1 = \frac{C}{c} \\
   C_3 = cC
   \]

7. Set \( C_1 \) and \( C_3 \) to the nearest standard values.

8. Recalculate \( C \), \( C^2 \), \( R \) and \( r^2 \):
   \[
   C = \sqrt{C_1C_3} \\
   c^2 = \frac{C_3}{C_1} \\
   R = \frac{1}{(\alpha_0C)} \\
   r^2 = \frac{1 + \sqrt{1 + 4Q_p^2(1 + c^2)(K - 1)}}{2 \cdot Q_p \cdot (1 + c^2) / c}^2
   \]

9. Calculate the resistors:
   \[
   \begin{align*}
   R_4 &= \frac{R}{r} \\
   R_5 &= rR
   \end{align*}
   \]

The component sensitivity formulas are in the table below. The sensitivities formulas are in the table below. The sensitivities to \( \alpha_i = K \) are a measure of this biquad’s sensitivity to the op amp group delay (reference [5]). To evaluate this biquad is sensitivity performance, use the method in reference [6].
4 KRC Highpass Biquad Parasitic Compensation

To pre-distort this biquad, and compensate for the [parasitic] non-inverting input capacitance of the op amp ($C_{ni}$), do the following (see Appendix A for the derivation of the formulas):

1. Start the iterations by ignoring the parasitics: $\tau^2 = 0$, $\tau_2 = 0$

2. Estimate the pre-distorted values of $\omega_p$ and $Q_p$ ($\omega_{p(pd)}$ and $Q_{p(pd)}$) that will compensate for $\tau_{oa}$ and $C_{ni}$:

$$\frac{\omega_{p(pd)}}{\omega_{p(nom)}} = \sqrt{1 - \tau_2^2 \omega_{p(nom)}^2}$$

$$Q_{p(pd)} = Q_{p(nom)} \frac{\omega_{p(pd)}}{\omega_{p(nom)}} \tau_2 \omega_{p(pd)}$$

Where $\omega_{p(nom)}$ and $Q_{p(nom)}$ are the nominal values of $\omega_p$ and $Q_p$.

3. Recalculate the resistors and capacitors using $\omega_{p(pd)}$ and $Q_{p(pd)}$:

$$\frac{1}{\omega_{p(pd)}} = \frac{1}{\omega_{p(pd)}} R_4 R_5 C_1 C_3$$

$$\frac{1}{\omega_{p(pd)}Q_{p(pd)}} = R_3 C_1 + R_5 C_3 - R_4 C_3 (K - 1)$$

Section 5 accomplishes this by recalculating $R$ and $r^2$, then $R_4$ and $R_5$:

$$R = \frac{1}{(\omega_{p(pd)}C)}$$

$$r^2 = \frac{1 + \sqrt{1 + 4Q_{p(pd)}^2 (1 + c^2) (K - 1)}}{2 \cdot Q_{p(pd)}(1 + c^2)/c}$$

$$R_4 = \frac{R}{r}$$

$$R_5 = rR$$

<table>
<thead>
<tr>
<th>$\alpha_i$</th>
<th>$S^{H_i}_{\alpha_i}$</th>
<th>$S^{\omega_p}_{\alpha_i}$</th>
<th>$S^{Q_p}_{\alpha_i}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_1$</td>
<td>0</td>
<td>$-\frac{1}{2}$</td>
<td>$-\left(Q_p \cdot \frac{r}{c} - \frac{1}{2}\right)$</td>
</tr>
<tr>
<td>$C_3$</td>
<td>0</td>
<td>$-\frac{1}{2}$</td>
<td>$\left(Q_p \cdot \frac{r}{c} + \frac{1}{2}\right)$</td>
</tr>
<tr>
<td>$R_4$</td>
<td>0</td>
<td>$-\frac{1}{2}$</td>
<td>$\left((K - 1) \cdot Q_p \cdot \frac{c}{r} + \frac{1}{2}\right)$</td>
</tr>
<tr>
<td>$R_5$</td>
<td>0</td>
<td>$-\frac{1}{2}$</td>
<td>$\left((K - 1) \cdot Q_p \cdot \frac{c}{r} - \frac{1}{2}\right)$</td>
</tr>
<tr>
<td>$R_f$</td>
<td>$\frac{K - 1}{K}$</td>
<td>0</td>
<td>$\left((K - 1) \cdot Q_p \cdot \frac{c}{r}\right)$</td>
</tr>
<tr>
<td>$R_{r_f}$</td>
<td>$-\frac{K - 1}{K}$</td>
<td>0</td>
<td>$\left((K - 1) \cdot Q_p \cdot \frac{c}{r}\right)$</td>
</tr>
<tr>
<td>$K$</td>
<td>1</td>
<td>0</td>
<td>$\left(K \cdot Q_p \cdot \frac{c}{r}\right)$</td>
</tr>
</tbody>
</table>
4. Calculate the resulting parasitic correction factors:
\[ r^2 = R_4 C_n t_4^2 = K t_{oa} R_4 R_6 (C_1 + C_3) C_{ni} \]

5. Calculate the resulting filter response parameters \( \omega_p \) and \( Q_p \)
\[
\omega_p = \frac{\omega_{p(pd)}}{\sqrt{1 + \tau_4^2 \omega_{p(pd)}^2}} \\
Q_p = \frac{Q_{p(pd)}}{\sqrt{\frac{\omega_p}{\omega_{p(pd)}} + Q_{p(pd)} \tau_2^2 \omega_p}}
\]

6. Repeat steps 2-5 until:
\[
\omega_p = \omega_{p(nom)} \\
Q_p = Q_{p(nom)}
\]

7. Estimate the high frequency gain:
\[
H_\infty = K \frac{1}{1 + \tau_4^2 \omega_{p(pd)}^2}
\]
If this reduces the gain too much, then repartition the gain.

5 Design Example

The circuit shown in Figure 2 is a 3rd-order Butterworth highpass filter. Section A is a buffered single pole section, and Section B is a highpass biquad. Use a voltage source with low output impedance, such as the CLC111 buffer, for \( V_{in} \):

The nominal filter specifications are:
\[
\begin{align*}
    f_c &= 50\text{MHz} -- (\text{passband edge frequency}) \\
    f_s &= 10\text{MHz} -- (\text{stopband edge frequency}) \\
    f_{hi} &= 200\text{MHz} -- (\text{highest signal frequency}) \\
    A_p &= 3.0\text{dB} -- (\text{maximum passband ripple}) \\
    A_s &= 40\text{dB} -- (\text{minimum stopband attenuation}) \\
    H_\infty &= 0\text{dB} -- (\text{passband voltage gain})
\end{align*}
\]

![Figure 2. Highpass Filter](image-url)
The 3rd-order Butterworth filter [1-4] meets our specifications. The pole frequencies and quality factors are:

<table>
<thead>
<tr>
<th>Section</th>
<th>A</th>
<th>B</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \omega_p / 2\pi [\text{MHz}] )</td>
<td>50.00</td>
<td>50.00</td>
</tr>
<tr>
<td>( Q_p [\text{]} )</td>
<td>–</td>
<td>1.000</td>
</tr>
</tbody>
</table>

**Overall Design**
1. Restrict the resistor and capacitor ratios to:
2. \( 0.1 \leq c^2, r^2 \leq 10 \)
3. Use 1% resistors (chip metal film, 1206 SMD)
4. Use 5% capacitors (ceramic chip, 1206 SMD)
5. Use standard resistor and capacitor values

**Section A Design and Pre-distortion:**
1. Use the CLC111. This is a close-loop buffer.
   - \( f_{3dB} = 800\text{MHz} > f_{H} = 200\text{MHz} \)
   - \( f_{3dB} = 800\text{MHz} > 10f_c = 500\text{MHz} \)
   - \( SR = 3500\text{V}/\mu\text{s}, \text{while a } 200\text{MHz, } 2V_{pp} \text{ sinusoid requires more than } 100\text{V}/\mu\text{s} \)
   - \( \tau_{oa} \approx 0.28\text{ns at } 10\text{MHz} \)
   - \( C_{n(111)} = 1.3\text{pF (input capacitance)} \)
2. Select \( R_{2A} \) for noise, distortion and to properly isolate the CLC111’s output and \( C_{1A} \). The pre-distorted value of \( R_{2A} \), that also compensates for \( C_{n(111)} \), is [5]:

\[
R_{2A} = \frac{1}{\left( \frac{\omega_p - \tau_{oa}}{C_{1A} + C_{n(111)}} \right)}
\]

The results are in the table below:
- The Initial Value column shows ideal values that ignore any parasitic effect
- The Adjusted Value column shows the component values that compensate for \( C_{n(111)} \) and CLC111 is group delay (\( \tau_{oa} \))
- The Standard Value column shows the nearest standard 1% resistors and 5% capacitors

<table>
<thead>
<tr>
<th>Component</th>
<th>Initial Value</th>
<th>Value Adjusted</th>
<th>Standard Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( C_{1A} )</td>
<td>30pF</td>
<td>30pF</td>
<td>30pF</td>
</tr>
<tr>
<td>( R_{2A} )</td>
<td>106Ω</td>
<td>92.8Ω</td>
<td>93.1Ω</td>
</tr>
<tr>
<td>( C_{n(111)} )</td>
<td>–</td>
<td>1.3pF</td>
<td>1.3pF</td>
</tr>
</tbody>
</table>
Design Example

Section B Design:

1. Since \( Q_p = 1.000 \), set \( K_B \) to 1.00

2. Use the CLC446. This is a current-feedback op amp
   - \( f_{3dB} = 400\text{MHz} > f_H = 200\text{MHz} \)
   - \( f_{3dB} < 10f_c = 500\text{MHz}; \) the design will be sensitive to the op amp group delay
   - \( SR = 2000\text{V/μs} > 1000\text{V/μs} \) (see Item #1 in "Section A Design")
   - \( t_{oa} \approx 0.56\text{ns at 10MHz} \)
   - \( C_{ni(446)} = 1.0\text{pF} \) (input capacitance)

3. Use the CLC446's recommended \( R_i \) at \( A_v = 1.0 \):
   - \( R_{IB} = 453Ω \)
   - Then leave \( R_{gB} \) open so that \( K_B = 1.00 \)

4. Initialize the resistor level:
   - \( R \approx 100Ω \)

5. Initialize the capacitor level, and the component ratios:
   - \( C = \frac{1}{2\pi(50.00\text{MHz})(100Ω)} = 31.83\text{pF} \)
   - \( c^2 = 0.1000 \)
   - \( r^2 = \max\{0.10,0.0826\} = 0.1000 \)

6. Recalculate \( C \) and initialize the capacitors:
   - \( C = 0.127 \quad C_{1B} \approx 89.3\text{pF} \quad C_{3B} \approx 11.3\text{pF} \)

7. Set the capacitors to the nearest standard values:
   - \( C_{1B} \approx 91\text{pF} \quad C_{3B} \approx 11\text{pF} \)

8. Recalculate the capacitor level and ratio, and the resistor level and ratio:

9. Calculate the resistors:
   - \( R_{4B} = 324Ω \quad R_{3B} = 31.2Ω \)

10. The sensitivities for this design are:

<table>
<thead>
<tr>
<th>( α_i )</th>
<th>( S_{\alpha_i}^H )</th>
<th>( S_{\alpha_i}^{ΩP} )</th>
<th>( S_{\alpha_i}^{ΩP} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>( C_{1B} )</td>
<td>0.00</td>
<td>-0.50</td>
<td>-0.39</td>
</tr>
<tr>
<td>( C_{3B} )</td>
<td>0.00</td>
<td>-0.50</td>
<td>0.39</td>
</tr>
<tr>
<td>( R_{4B} )</td>
<td>0.00</td>
<td>-0.50</td>
<td>0.50</td>
</tr>
<tr>
<td>( R_{3B} )</td>
<td>0.00</td>
<td>-0.50</td>
<td>-0.50</td>
</tr>
<tr>
<td>( R_B )</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
</tr>
<tr>
<td>( R_{gB} )</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
</tr>
<tr>
<td>( K )</td>
<td>1.00</td>
<td>0.00</td>
<td>1.12</td>
</tr>
</tbody>
</table>
Section B Pre-distortion:

1. The design gives these values:
   - $\omega_{p(nom)} = 2\pi(50.00\text{MHz})$
   - $Q_{p(nom)} = 1.000$
   - $K_B = 1.00$
   - $C_{1B} = 91\text{pF}$
   - $C_{3B} = 11\text{pF}$

2. Iteration 1 shows the initial design results. Iterations 2-4 pre-distort $R_{4B}$ and $R_{5B}$ to compensate for the CLC446’s group delay, and for $C_{ni(446)}$:

<table>
<thead>
<tr>
<th>Iteration #</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\omega_{p(pd)}/2\pi$</td>
<td>[MHz]</td>
<td>50.00</td>
<td>59.73</td>
<td>56.81</td>
</tr>
<tr>
<td>$Q_{p(pd)}$</td>
<td>[ ]</td>
<td>1.000</td>
<td>0.9320</td>
<td>0.9561</td>
</tr>
<tr>
<td>$R$</td>
<td>[Ω]</td>
<td>10.6</td>
<td>84.22</td>
<td>88.54</td>
</tr>
<tr>
<td>$r_2$</td>
<td>[Ω]</td>
<td>0.0962</td>
<td>0.1108</td>
<td>0.1053</td>
</tr>
<tr>
<td>$R_{4B}$</td>
<td>[Ω]</td>
<td>324.3</td>
<td>253.0</td>
<td>272.9</td>
</tr>
<tr>
<td>$R_{5B}$</td>
<td>[Ω]</td>
<td>31.21</td>
<td>28.03</td>
<td>28.73</td>
</tr>
<tr>
<td>$\tau_2$</td>
<td>[ns]</td>
<td>0.324</td>
<td>0.253</td>
<td>0.273</td>
</tr>
<tr>
<td>$\tau_4$</td>
<td>[ns]</td>
<td>1.741</td>
<td>1.511</td>
<td>1.575</td>
</tr>
<tr>
<td>$\omega_{p}/2\pi$</td>
<td>[MHz]</td>
<td>43.87</td>
<td>51.96</td>
<td>49.52</td>
</tr>
<tr>
<td>$Q_\omega$</td>
<td>[ ]</td>
<td>1.034</td>
<td>0.984</td>
<td>1.003</td>
</tr>
</tbody>
</table>

- The midband gain estimate is:
  - $\text{H}_\infty \approx 0.770 \text{[V/V]}$. Iteration 1 $\approx 0.759 \text{[V/V]}$. Iteration 4 (3)

- The simulations gave a lower value for $H_\infty$. Increasing $K$ could help overcome this loss, but would also increase the sensitivities.

3. The resulting components are:

<table>
<thead>
<tr>
<th>Component</th>
<th>Initial</th>
<th>Value Adjusted</th>
<th>Standard</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_{1B}$</td>
<td>91pF</td>
<td>91pF</td>
<td>91pF</td>
</tr>
<tr>
<td>$C_{3B}$</td>
<td>11pF</td>
<td>11pF</td>
<td>11pF</td>
</tr>
<tr>
<td>$C_{ni(446)}$</td>
<td>–</td>
<td>1.0pF</td>
<td>1.0pF</td>
</tr>
<tr>
<td>$R_{4B}$</td>
<td>324Ω</td>
<td>268Ω</td>
<td>267Ω</td>
</tr>
<tr>
<td>$R_{5B}$</td>
<td>31.2Ω</td>
<td>28.5Ω</td>
<td>28.7Ω</td>
</tr>
<tr>
<td>$R_{4B}$</td>
<td>453Ω</td>
<td>453Ω</td>
<td>453Ω</td>
</tr>
<tr>
<td>$R_{5B}$</td>
<td>–</td>
<td>–</td>
<td>–</td>
</tr>
</tbody>
</table>

Figure 3 and Figure 4 show simulated gains. The curve numbers are:

1. Ideal (Initial Design Values, $\tau_{oa} = 0$, $C_{ni} = 0$)
2. Without pre-distortion (Initial Design Values, $\tau_{oa} \neq 0$, $C_{ni} = 0$)
3. With pre-distortion (Pre-distorted Values, $\tau_{oa} \neq 0$, $C_{ni} = 0$)
6 SPICE Models

SPICE Models are available for most of Comlinear’s amplifiers. These models support nominal DC, AC, AC noise and transient simulations at room temperature.

We recommend simulating with Comlinear’s SPICE model to:

- Predict the op amp’s influence on filter response
- Support quicker design cycles

Include board and component parasitic models to obtain a more accurate prediction of the filter’s response.

To verify your simulations, we recommend bread-boarding your circuit.

7 Summary

This application report contains an easy to use design algorithm for a low sensitivities Sallen-Key highpass biquad. Designing for low $\omega_p$ and $Q_p$ sensitivities gives:

- Reduced filter variation over process, temperature and time
- High manufacturing yield
- Lower component cost

A low sensitivity design is not enough to produce high manufacturing yields. This Application Note shows how to compensate for the op amp bandwidth, and for the [parasitic] input capacitance of the op amp. This method also applies to any other component or board parasitics. The components must also have low enough tolerance and temperature coefficients.
Appendix A Derivation of Pre-distortion and Parasitic Capacitance Compensation Formulas

To pre-distort this filter, and compensate for the [parasitic] input capacitance of the op amp $C_{ni}$:

1. Use the method in reference [5] to include the op amp’s effect on the filter response. The result is:

$$\frac{V_0}{V_{in}} = \frac{H_0 \left( \frac{1}{\omega_p^2} \right) s^2}{1 + \left( \frac{1}{\omega_p Q_p} \right) s + \left( \frac{1}{\omega_p} \right) s^2} e^{-\tau_{oa} s}$$

where the op amp group delay ($\tau_{oa}$) is evaluated at the passband edge frequency ($f_c$), and:

$$\frac{1}{\omega_p Q_p} = R_5 C_1 + R_5 C_3 - R_4 C_3 (K - 1)$$

$$\frac{1}{\omega_p^2} = R_4 R_5 C_1 C_3 + K \tau_{oa} R_4 C_3$$

$$K = \frac{1 + R_4}{R_g}$$

$$H_0 = K$$

2. Since $C_{ni}$ is in parallel with $R_4$, replace $R_4$ with the parallel equivalent of $R_4$ and $C_{ni}$:

$$\frac{R_4}{1 + R_4 C_{ni} s}$$

$$\frac{V_0}{V_{in}} = \frac{H_0 \left( \frac{R_4 C_3 (R_5 C_1 + K \tau_{oa})}{1 + R_4 C_{ni} s} \right) s^2 e^{-\tau_{oa} s}}{1 + \left( \frac{R_4 C_3 (1 - K)}{1 + R_4 C_{ni} s} + R_5 (C_1 + C_3) \right) s + \left( \frac{R_4 C_3 (R_5 C_1 + K \tau_{oa})}{1 + R_4 C_{ni} s} \right) s^2}$$
3. After simplifying, we obtain:

\[
\frac{V_o}{V_{in}} = H_\infty \left( \frac{\omega_p}{\omega} \right)^2 e^{-\tau \omega^2} \cdot e^{-\tau \omega^2}
\]

where:

\[
\frac{1}{\omega_p Q_p} = t_1 + t_2
\]

\[
\tau_1 = R_5 C_1 + R_5 C_3 - R_4 C_3 (K - 1)
\]

\[
\tau_2 = R_4 C_{ni}
\]

\[
\frac{1}{\omega_p^2} = \tau_3^2 + \tau_4^2
\]

\[
\tau_3^2 = R_4 R_5 C_1 C_3
\]

\[
\tau_4^2 = K \tau_{on} R_4 C_3 + R_4 R_5 (C_1 + C_3) C_{ni}
\]

\[
K = \frac{1 + R_4}{R_4}
\]

\[
H_\infty = \frac{K \cdot \left( \tau_3^2 \right)}{\left( \tau_3^2 + \tau_4^2 \right)}
\]
Appendix B Bibliography

5. *OA-21 Component Pre-Distortion for Sallen Key Filters Application Report* (SNOA369)
7. *OA-28 Low-Sensitivity, Bandpass Filter Design with Tuning Method Application Report* (SNOA373)

Note: The circuits included in this application note have been tested with Texas Instruments parts that may have been obsoleted and/or replaced with newer products. Please refer to the CLC to LMH conversion table to find the appropriate replacement part for the obsolete device.
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